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# SYSTEM ANALYSIS OF ULTRA-WIDEBAND INSTRUMENTATION RADARS: IMPULSE VS. STEPPED-CHIRP APPROACHES

James Ralston



November 1993



*Prepared for*
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INSTITUTE FOR DEFENSE ANALYSES
1801 N. Beauregard Street, Alexandria, Virginia 22311-1772

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As part of an ongoing effort to determine the utility of ultra-wideband (UWB) radar systems for military application, an experimental program was developed and executed to collect terrain clutter data using high resolution waveforms in the ultra high frequency (UHF) spectral region. Two approaches to the design of the radar instrumentation to be used to collect these data were considered: an impulse system with a nominal 1 ns pulse duration and a "conventional" stepped-chirp instrumentation radar covering the same frequency range. A novel feature of the program was the use of a scanned linear aperture to simulate the use of a large, ideally weighted real aperture antenna system. In this paper, the theoretical analysis done to predict and compare the performance expected from either system approach is presented in terms of the noise-equivalent reflectivity of the clutter measurement system, the time to collect data, and the impact of the linearly scanned aperture on sensitivity, angular resolution, and data collection time.

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# PREFACE

This work was performed for the Advanced Research Projects Agency (ARPA) to assist in planning, developing, and evaluating the Radar System Development program, specifically in the area of ultra-wideband radar (UWB) technology. The author wishes to thank Dominick Giglio, George Ruck, Michael Tuley, Karl Davis, Arthur Krinitz, and Guillermo Loubriel for their many helpful observations.

## ABSTRACT

As part of an ongoing effort to determine the utility of ultra-wideband (UWB) radar systems for military application, an experimental program was developed and executed to collect terrain clutter data using high resolution waveforms in the ultra high frequency (UHF) spectral region. Two approaches to the design of the radar instrumentation to be used to collect these data were considered: an impulse system with a nominal 1 ns pulse duration and a "conventional" stepped-chirp instrumentation radar covering the same frequency range. A novel feature of the program was the use of a scanned linear aperture to simulate the use of a large, ideally weighted real aperture antenna system. In this paper, the theoretical analysis done to predict and compare the performance expected from either system approach is presented in terms of the noise-equivalent reflectivity of the clutter measurement system, the time to collect data, and the impact of the linearly scanned aperture on sensitivity, angular resolution, and data collection time.

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# ACRONYMS AND ABBREVIATIONS

| | |
|---|---|
| ARPA | Advanced Research Projects Agency |
| C/N | clutter-to-noise (ratio) |
| DARPA | Defense Advanced Research Projects Agency |
| dB | decibel |
| dBW | decibels above one watt |
| GHz | gigahertz |
| Hz | hertz |
| km | kilometer |
| kW | kilowatt |
| MHz | megahertz |
| ns | nanosecond |
| OSD | Office of the Secretary of Defense |
| PRF | pulse repetition frequency |
| RCS | radar cross-section |
| RFI | radio frequency interference |
| S/C | signal-to-clutter (ratio) |
| S/N | signal-to-noise (ratio) |
| TEM | transverse electromagnetic |
| UHF | ultra high frequency |
| UWB | ultra-wideband |
| W | watt |

# I. INTRODUCTION

## A. ULTRA-WIDEBAND (UWB) RADAR PROGRAM HISTORY

The Balanced Technology Initiative Office of the Office of the Secretary of Defense (OSD) began the UWB Technology program in 1990, with a panel study conducted under Defense Advanced Research Projects Agency (DARPA) auspices. This panel reviewed existing work in the field and developed a list of potential applications (Ref. 1). The technology and data required to support these applications were prioritized, and several possible areas of research were recommended. In particular, the panel suggested that point designs for UWB radar in several important applications be analyzed, that gaps in the phenomenological data base be filled, and that some effort be directed toward the development of key components.

For military radars, one of the most important aspects of UWB waveforms is the potential for enhancing signal-to-clutter (S/C) ratio for airborne targets. Tactical radars, in particular, are more often limited by clutter than by system receiver noise. This is especially true when trying to detect the types of low-altitude pop-up targets that may pose threats to mobile forces on the march. The rationale (Ref. 1) for evaluating UWB waveforms was that the high range resolution would reduce clutter cell area directly, while the low frequency content of the waveform could enhance scattering, particularly from targets that had been designed for low radar signature in the microwave bands. In addition, it was anticipated that clutter reflectivity should be moderate for low frequency waveforms, at least in some types of terrain. Recent work (Ref. 2) reported elsewhere, however, suggests that clutter reflectivity does not decrease monotonically with radar frequency in all cases.

Because impulse radars are able to combine high range resolution with low frequency spectral content, their potential for achieving high S/C is evident. On the other hand, it is well known that the amplitude distribution of clutter return can show strong variations with the size of the clutter resolution cell. Thus, the radar false alarm rate, which is the real determinant of detection performance in clutter, is not necessarily proportional to clutter cell area and may turn out to be only weakly dependent on resolution. Because existing data did not extend to the critical case of low frequency, high resolution, and low grazing angle, DARPA determined that a specifically focused clutter measurement program was required.

To afford an opportunity to develop and test UWB technology, it was desired that these measurements employ an impulse implementation of a UWB radar for at least some of the data collection. At the same time, it was recognized that there would be limitations in the data collection rate available from existing impulse radar technology and that it would be risky to depend solely on an untested impulse system for data collection in a remote field environment. For this reason, a system analysis of the expected performance of two specific implementations of a UWB radar—one employing impulse, the other stepped-chirp waveforms—was performed.

The scope and results of this measurement program are described in the following paper.

## B. WAVEFORM AND ANTENNA SELECTION

First of all, it was recognized that the measurement program would be highly limited in the scope of terrain examined and in duration. In addition, the types of impulse radar systems that could be readily engineered were generally limited to a single waveform. Thus, it was necessary to identify a single measurement waveform. Several factors were considered in the selection of this waveform, and these factors reflected the motivating military application and the practical instrumentation constraints. To achieve the possible benefits of low frequency and to be compatible with current impulse instrumentation receiver technology, it was desired that the dominant signal spectrum be well below 1 GHz.

Antenna sizing was another consideration. Regardless of waveform, good clutter performance requires low antenna sidelobes, which, in turn, require a minimum aperture of 6 to 10 wavelengths. To avoid placing undue requirements on antenna size, 300 MHz was used as a working lower limit on signal spectrum. Even at that, antenna apertures of 6 to 10 metres may be necessary, and this will present some interesting design challenges if such waveforms are to be used for mobile air defense systems.

### 1. Impulse Waveform Model

This work was directed at developing an analytic tool for guiding the development of UWB instrumentation radars (employing either impulse or chirped waveforms) for clutter measurement. Consequently, it was necessary to define an analytic model of impulse waveforms that was sufficiently general to allow system tradeoffs and to reflect the characteristics of the as-built impulse source. Several waveforms were investigated as paradigms for impulse radiators, and the one that appeared most appropriate for this purpose was a band-limited variation of the single-cycle sinewave, called the *monocycle* (Ref. 3). The monocycle and its associated energy spectral density are shown in Figure 1. The shape of the wave is defined by a single parameter, its duration, $\tau$. This is equal to the period of the generating sinewave. Thus, it is convenient to normalize time to units of $\tau$ and express frequency in units of reciprocal periods, $1/\tau$. For example, if the period indicated in Figure 1a is a nanosecond, then the frequency unit in Figure 1b is GHz. The energy spectrum shown in Figure 1b can be expressed analytically as

$$P_s(f) = \left[ \frac{\tau}{\pi} \bullet \frac{\sin(\pi f \tau)}{1-(f\tau)^2} \right]^2 \quad . \tag{1}$$

From Eq. (1), it is straightforward to derive expressions relating $\tau$ to the peak of the energy spectrum and the lower and upper 3-dB (half power) points in the spectrum, denoted $f_1$ and $f_2$, respectively:

$$\begin{aligned} f_{peak} &= 0.84/\tau \\ f_1 &= 0.411/\tau \\ f_2 &= 1.31/\tau \quad . \end{aligned} \tag{2}$$

The perfect monocycle illustrated in Figure 1a is only possible given unlimited bandwidth. Assuming that the waveform is sharply limited to the passband from $f_1$ to $f_2$, the resulting time sidelobes give it the shape shown in the figure for the band-limited case, which is typical of impulse UWB waveforms radiated by many real sources. For an ideal sinewave, the energy in a single cycle is $\tau \cdot P_{peak}/2$. For this model, an additional loss of 1 dB (factor of 0.8) is taken to account for band limiting in an imperfect matched filter. Thus, the energy transmitted in a single pulse is given by $E_t = 0.4 \cdot \tau \cdot P_{peak}$.

## 2. Antenna Considerations

When a wideband radar system is used in practical target detection, it is desired to obtain the maximum antenna performance across the frequency spectrum. Thus, it is accepted that the antenna gain will be limited by aperture diffraction at each frequency and, as a result, will vary over the frequency band. The angular resolution of the resulting system will be highly dispersive. For instrumentation purposes, however, in which we

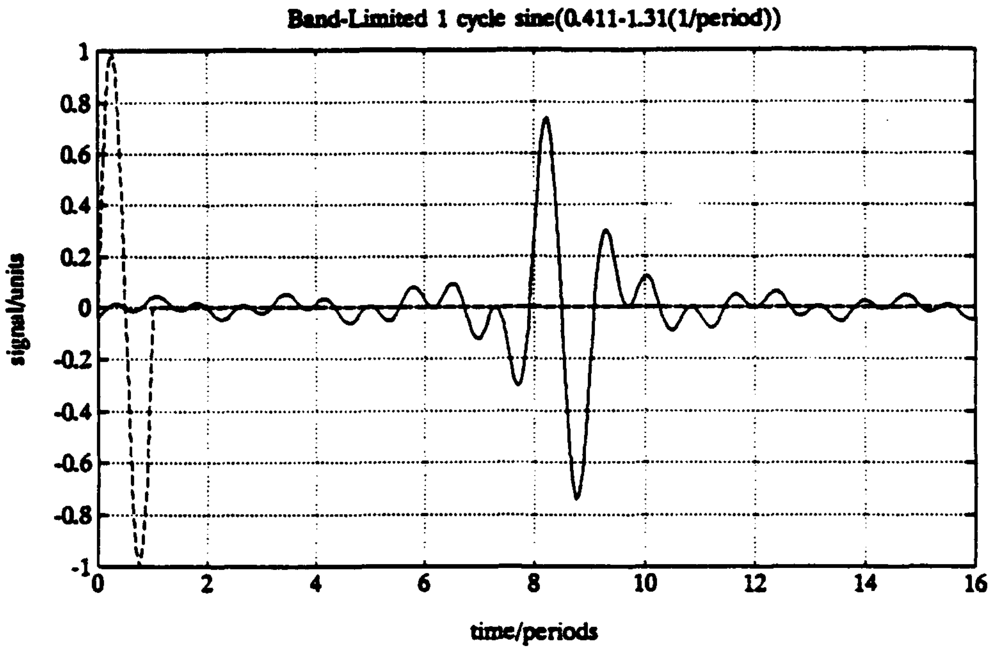


**Figure 1a. The Single-Cycle Sinewave, or *Monocycle* Waveform**

***(The dashed line shows an ideal monocycle of unit amplitude; the solid line is the same waveform after band limiting. See Figure 1b.)***

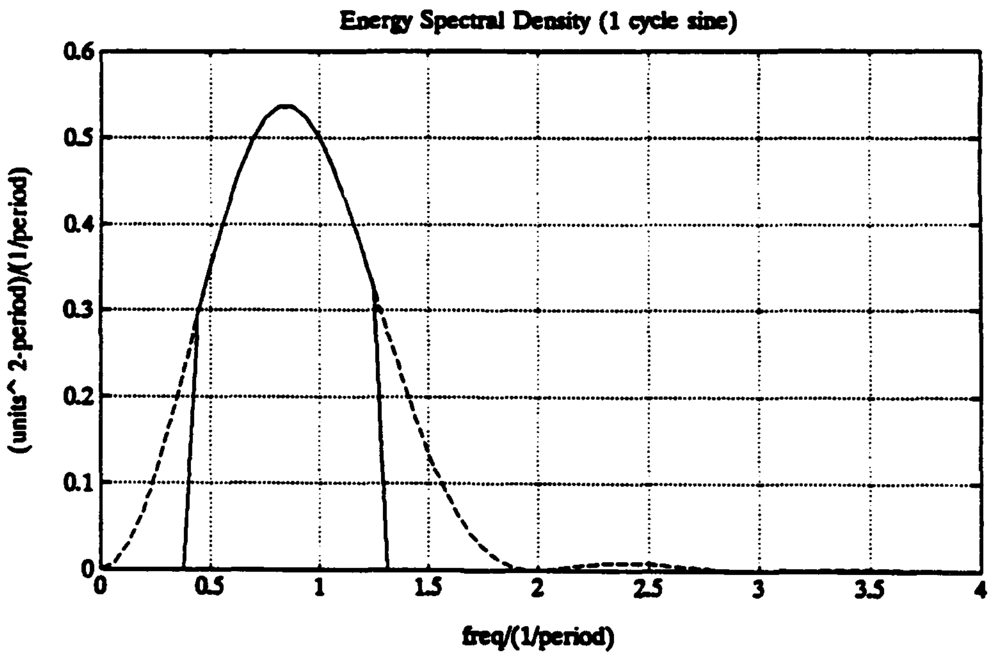


**Figure 1b. The Energy Spectral Density of the Monocycle Waveform**

***(The dashed line is the energy spectrum of the ideal monocycle. The solid line represents the same spectrum band-limited near the half-power points. The resolution of the Fourier routine employed was not sufficient to match the exact half-power points.)***

attempt to measure clutter reflectance by normalizing clutter radar cross-section (RCS) by the area of the clutter cell, frequency-dependent angular resolution will lead to highly ambiguous results. For that reason, an antenna that would maintain relatively constant beamwidth and gain over the frequency band of our signal was needed.

Another requirement was that the antenna aperture be large enough to provide reasonably high angle resolution and good sidelobe control at the lowest frequency used and still be transportable over unpaved roads. Both of these requirements were met by a synthetic aperture antenna obtained by mechanically scanning a small transverse electromagnetic (TEM) horn element over a spatial window of up to 20 metres. Constant synthetic beamwidth over the passband was obtained by Fourier transform of the received waveforms to the frequency domain and by appropriately weighting each frequency component as a function of element position before forming the synthetic beam. In this way, effective linear aperture could be scaled with wavelength to maintain constant beam characteristics in the azimuthal plane.

# II. RADAR SYSTEM ANALYSIS

## A. IMPULSE UWB RADAR

### 1. Impulse Radar Equation

The familiar radar equation essential to conventional radar system analysis is given by

$$\frac{S}{N} = \frac{E_t G A_e \sigma}{(4\pi)^2 R^4 kTLF} \quad , \tag{3}$$

where $E_t$ is the energy of the pulse, G the antenna gain, $A_e$ the effective antenna receive aperture, $\sigma$ the target RCS, R the target's range, kT the standard noise power density of 4 E -21 watts/Hz, L the system losses, and F the receiver noise figure. This equation assumes that the receiver is matched to the waveform. **Note:** This equation is only valid for narrowband waveforms.

There have been a number of attempts to derive an equivalently useful expression for UWB radar systems that can relate signal-to-noise (S/N) performance to waveform and radar parameters in the same way (Ref. 1). The approach chosen here is based on the assumptions that the signal power spectrum is flat over its passband (this is only an approximation to the power spectrum of Figure 1b) and that the antenna's gain and beamwidth are independent of frequency over the passband. In effect, this implies that the antenna aperture is efficiently used only at the lowest frequency and is degraded with increasing frequency. That is,

$$\begin{aligned} A_e &= A_0 \frac{\lambda^2}{\lambda_1^2} = A_0 \frac{f_1^2}{f^2} \\ G &= 4\pi \frac{A_e}{\lambda^2} = 4\pi \frac{A_0}{\lambda_1^2} = constant \\ \theta &= 1.3\lambda_1 / D \quad , \end{aligned} \tag{4}$$

where $A_0$ is the maximum value of the antenna effective aperture (only realized at the lowest frequency), $\lambda_1$ corresponds to the low-frequency limit of the signal passband, $\theta$ is the 3-dB antenna beamdwidth, and D is the antenna's horizontal dimension. A potential realization of such an antenna is discussed in Reference 1. With these assumptions, Eq. (3) can be integrated between the limits $f_1$ and $f_2$ to obtain a "wideband radar equation":

$$\frac{S}{N}=\frac{E_t G A_0 \sigma}{(4\pi)^2 R^4 kTLF}\bullet\frac{f_1}{f_2} \quad . \tag{5}$$

This is identical to the narrow band equation (3) with the addition of the factor $f_1/f_2$. Because this factor is always < 1, it can be thought of as an equivalent "wideband loss" resulting from constraining the antenna performance to its value at the lowest frequency.

In this case, the signal is that of clutter, with an RCS given by $\sigma_c = \sigma_0 \bullet A_c$, where $\sigma_0$ is the clutter reflectivity that will ultimately be measured and $A_c$ is the clutter cell area. The cell area is given by

$$A_c = R\theta\Delta r = R\bullet\frac{1.3\lambda_1}{D}\bullet\frac{c\tau}{2} \quad . \tag{6}$$

Substitution into Eq. (5) yields

$$\frac{C}{N}=\frac{E_t G A_0}{(4\pi)^2 R^3 kTLF}\bullet\sigma_0\bullet\frac{c\tau}{2}\bullet\frac{1.3\lambda_1}{D}\bullet\frac{f_1}{f_2} \tag{7}$$

for the effective single-pulse clutter-to-noise (C/N) ratio of an impulse radar using a real-aperture antenna element of dimension, D. Now, let this element be scanned over $N_s$ successive beam positions at an interval $l$. Assuming full coherent processing of the successive pulses and that $N_s >> 1$, the effective post-processing waveform energy is $E_{ts} = N_s \bullet E_t$ and effective post-processing beamwidth is $\theta_s = 1.3\lambda_1/N_s \bullet l$. Substituting the post-processing values for the single-pulse values in Eq. (7) gives:

$$\frac{C}{N}=\frac{E_t G A_0}{(4\pi)^2 R^3 kTLF}\bullet\sigma_0\bullet\frac{c\tau}{2}\bullet\frac{1.3\lambda_1}{\ell}\bullet\frac{f_1}{f_2} \quad . \tag{8}$$

Comparing Eqs. (8) and (7) shows that the ratio of C/N for the synthetically scanned antenna system and the single real element is

$$\frac{(C/N)_{synth}}{(C/N)_{real}}=\frac{D}{\ell} \quad . \tag{9}$$

This is independent of $N_s$. In other words, increasing the length of the scanned aperture leads to a reduction in $A_c$ that balances the increase in total waveform energy transmitted. If the scan step interval is equal to the antenna's effective aperture dimension, the resulting C/N is not changed. Thus, there is no improvement in the floor of measurement sensitivity. However, there is a reduction in cell area, allowing the measurements of clutter statistics on a finer scale. If many pulses are coherently averaged per element dimension, corresponding to averaging many pulses in each position of the scanner, an effective C/N processing gain will be obtained.

## 2. Impulse Noise-Equivalent Clutter Reflectivity

To obtain a useful design equation for predicting the performance of a scanned-aperture impulse radar system in measuring clutter, several assumptions are made. First, the scanner step size is one half the wavelength of the highest frequency component ($l = \lambda_2/2$). This is a very stringent condition that guarantees grating lobes will not occur even if the synthetic beam is scanned to near endfire. Another assumption is that the monocycle waveform is used. This allows the pulse energy, frequency, and wavelength values that occur to be expressed in terms of the pulsewidth, $\tau$, and the peak power. In addition, Eq. (4) is used to characterize the antenna in terms of its gain, G. Solving for the noise-equivalent clutter reflectivity, or $NE\sigma_0$, the following is obtained:

$$NE\sigma_0 = \frac{(4\pi)^3 kT(0.411)^2}{1.3c^3(0.4)} \bullet \frac{R^3 LF}{P_{peak}\tau^4 G^2 N_p} \quad , \tag{10}$$

where $N_p$ is the number of pulses coherently averaged at each position of the scanner. The factor 0.4 introduced in the denominator reflects the fact that the effective power over the pulsewidth, $\tau$, is 0.5 times the peak value and that there is an additional 1 dB (factor of 0.8) mismatch loss due to band-limiting. Consolidating constants, using convenient units, and expressing the results in dB, Eq. (10) can be rewritten as follows:

$$\begin{aligned} NE\sigma_0 &= 19.8dB + 30\log(R/km) + LF - 2G \\ &\quad -10\log P_{peak}(W) - 40\log\tau(ns) \\ &\quad -10\log N_p \quad . \end{aligned} \tag{11}$$

Eq. (11) can now be applied to estimate the performance of the impulse instrumentation system in collecting clutter data.

Despite its remote location, the environment of the clutter measurements showed a high level of radio frequency interference (RFI) in the frequency bands of interest. Measurements (Ref. 4) indicated that the effective level of RFI power in the receiver was 17 dB above thermal noise. For that reason, a total of system losses plus noise figure (LF) of 20 dB was assumed. Antenna element gain was assumed approximately constant at 8 dB across the band. Pulse monocycle duration was adjusted to 1.3 ns to place the bulk of the pulse energy between 300 and 1000 MHz. The peak power of the pulser was estimated to be 40 kW. Substitution into Eq. (10) yields the dependence of $NE\sigma_0$ on range shown in Figure 2. The line plotted corresponds to $N_p = 1$. The difference between the single-pulse sensitivity and the required noise floor at any range represents the processing gain required. In this example, a need to make clutter measurements as low as –30 dB with a desired C/N ratio of 10 dB is anticipated. Thus, the desired measurement floor is –40 dB.

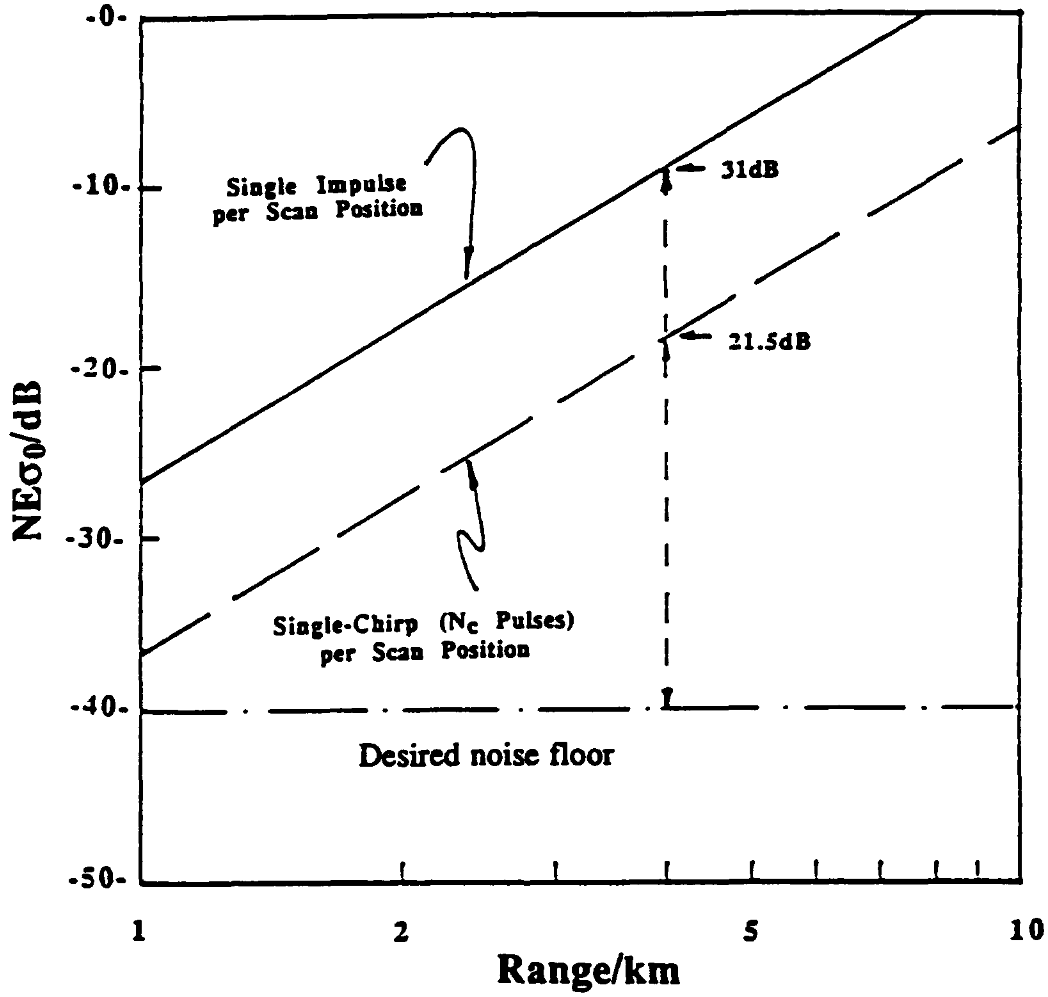


**Figure 2. Predicted Noise-Equivalent Clutter Reflectivities as a Function of Range for the Specific Impulse and Stepped-Chirp Radar Systems Considered for Clutter Data Collection**

***[Sensitivity corresponds to a single pulse for the impulse system and a single chirp (one pulse per frequency step) for the chirped system, at each position of the antenna scanner. For a given range (4 km chosen as an example) the difference between the single-waveform sensitivity and the required noise measurement floor (–40 dB chosen as an example) defines the number of waveforms required to be coherently processed at each scanner position. The required processing gain is 21.5 dB (141 complete chirps) for the chirped system and 31 dB (1260 pulses) for the impulse system.]***

## B. CHIRPED UWB RADAR

In the stepped-chirp system, the frequency interval $f_1$–$f_2$ is filled by $N_c$ single-frequency pulses spaced by $\Delta f$. The resulting chirp bandwidth is $B_c = N_c \bullet \Delta f$. The range resolution of this waveform is $\Delta r = c/2B_c$. To avoid aliasing, the uncompressed pulse-width (the pulsewidth of the individual single-frequency pulses) and the frequency step interval must satisfy $\tau_p \leq 1/\Delta f$. In the stepped-chirp system, the compressed pulsewidth, $\tau_c$, corresponds to the monocycle pulsewidth, $\tau$. A stepped-chirp radar system can be used to excite the same scanned-aperture antenna system that is used by the impulse radar. In this case, however, a minimum of $N_c$ pulses must be processed at each scanner position to transmit a complete waveform. After processing one pulse at each point in frequency and position space, for a total of $N_s \bullet N_c$ pulses, the C/N is

$$\frac{C}{N} = \frac{PG^2\lambda_1^2\sigma_0\theta\Delta r}{(4\pi)^3R^3kTB_nLF} \bullet N_sN_c \bullet \frac{f_1}{f_2}$$

$$= \frac{PG^2c^4(1.3)\sigma_0}{2(4\pi)^3R^3kTLFB_n\Delta f\ell f_1^2f_2} \quad . \tag{12}$$

Using Eq. (2) to relate the frequency limits to the compressed pulsewidth, the C/N values of the stepped-chirp and impulse waveforms can be compared by taking the ratio of Eqs. (12) and (8). The result is

$$\frac{(C/N)_c}{(C/N)_i} = \frac{N_cE_{tc}}{E_{ti}} \bullet \frac{m_i}{m_c} \quad , \tag{13}$$

where $E_{tc}$ is the energy of a single pulse. The product of $E_{tc} \bullet N_c$, therefore, is the total energy of a single chirped waveform. The "m" parameters are the filter mismatch factors for the impulse and chirped waveforms. Recall that a mismatch of 1 dB for the impulse waveform was assumed. The mismatch for chirped systems is $m_c = B_n \bullet \tau_p$. To maintain good transient response and minimize aliasing, it is conventional for stepped-chirp systems to operate at a wider system bandwidth than $1/\tau_p$. Factors of order 10 for $m_c$ are typical. Eq. (12) shows that, mismatch factors aside, the ratio of processed C/N values per waveform is simply the ratio of waveform energy, other system factors and losses being equal. For a stepped-chirp system, the noise-equivalent clutter reflectivity is given by

$$NE\sigma_0 = \frac{(4\pi)^3kT(0.411)^2}{1.3c^3} \bullet \frac{R^3LFB_n}{P\tau_pG^2\tau_c^2} \quad . \tag{14}$$

Expressed in dB, this becomes

$$\begin{aligned} NE\sigma_0 = {} & 15.8dB + 30\log R(km) + LF + 10\log B_n(GHz) \\ & -10\log P(W) - 10\log \tau_p(ns) - 20\log \tau_c(ns) \\ & -2G - 10\log N_p \quad . \end{aligned} \tag{15}$$

This is applied to the same data collection problem posited for the impulse radar. LF equals 20 dB, again reflecting RFI, and $\tau_c$ equals 1.3 ns, corresponding to the impulse case. If the number of frequency steps in the chirped waveform, $N_c$, is 1024, a convenient value for chirp programming, then the length of the single-frequency pulse, $\tau_p$, is 1331 ns. A typical value of system bandwidth, which defines the noise bandwidth, would be 0.01 GHz. Element gain is 8 dB, as above. Transmitter power is 2W (3 dBW). Although this represents a much smaller peak power than the 40 kW estimated for the impulse system, the pulse duration is 1024 times longer and there are 1024 times as many pulses in the waveform. Thus, despite the 11 dB additional mismatch loss, the $NE\sigma_0$ performance of the chirped system is much better than that of the impulse system, as shown in Figure 2.

## C. COMPARATIVE PERFORMANCE

Because both the impulse system and the stepped-chirp system are free to trade data collection time against measurement noise floor, the most meaningful comparison is based on estimated data collection time for a common level of sensitivity. To obtain a measurement noise floor of –40 dB at a range of 4 km, the impulse system requires a coherent processing gain of 31 dB, corresponding to averaging 1250 pulses at each scanner position. To achieve the same sensitivity, the stepped-chirp system needs 21.5 dB of gain, corresponding to averaging 141 pulses at each frequency for each scanner position. For the particular aperture scanner used in these measurements, the time required to traverse the aperture is 12 seconds per metre for motion, plus 4.16 seconds for each stop and start. The time to complete a single scan is

$$T_{scan} = (12\sec/m)L_s + (4.16\sec/pos)N_s + \frac{N_p N_s N_c}{f_p} \quad , \tag{16}$$

where $L_s$ is the scan length in metres (20 metres for both), $N_s$ is the number of steps in one scan (256 for both), $N_p$ is the number of pulses per frequency at each scanner position (about 1250 for the impulse system, 140 for the chirp system), $N_c$ is the number of single-frequency pulses in one chirp (1 for the impulse radar, 1024 for the chirp radar), and $f_p$ is the radar pulse repetition frequency (PRF) (20 Hz for the impulse system, 55 kHz for the chirp system). The time per scan is approximately 33 minutes for the chirped system and 290 minutes for the impulse system.

The performance differential becomes greater if we compute the time to collect a unit area of clutter data. The swath of the impulse system, based on the memory limitations of the digitizer used to collect the data, is 38 metres. The corresponding swath of the stepped-chirp waveform, based on the uncompressed pulsewidth of 1331 ns, is 200 metres. In combination, these two factors indicate that the chirped system would have a net advantage of almost a factor of 50 in data collection throughput.

This comparison is between two specific examples of a UWB radar system and, as such, is somewhat unfair to the impulse system since the chirped system used as a benchmark represents a third-generation system design based on a mature base of available components and several years of system improvement. By comparison, impulse systems are in their infancy. The PRF of 20 Hz for the impulse system results from hardware limitations in the particular digitizer that is used and is not an inherent constraint. Higher power pulse generators could also be used. However, although significant improvements in the throughput of impulse radars are possible, there is little reason to expect that impulse systems will do better than approach the performance of chirped systems in the instrumentation role.

# III. CONCLUSIONS

This analysis has established that an impulse radar system working around proven, fieldable components could be built and would be able to collect at least a portion of the desired UWB clutter data. The data collection speed of this system, however, would be limited by the low average power of the system, which would be limited by the available peak power of the pulse source chosen and by the low PRF mandated by the available digitizing electronics. Therefore, to achieve the program goals (i.e., maximize data collection and minimize risk), it was decided to use both of the radar systems analyzed in the paper.

# REFERENCES


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